APPLICATION FOR UNITED STATES LETTERS PATENT

FOR

METHOD AND APPARATUS FOR SEMI-BLIND COMMUNICATION CHANNEL ESTIMATION

Inventors:

Behrouz Farhang-Boroujeni
Francois Po Shin Chin
Chin Keong Ho

Express Mail Label No. EL743033932US
Date of Deposit: February 6, 2001

METHOD AND APPARATUS FOR SEMI-BLIND COMMUNICATION CHANNEL ESTIMATION

Field of the Invention

5

The present invention relates to multi-carrier communication systems, and more particularly to channel estimation in multi-carrier communication systems that employ orthogonal frequency division multiplexing (OFDM).

10   Background of the Invention

OFDM is a multi-carrier transmission technique, which divides available frequency spectrum of a communication channel into many carriers, often referred to as sub-carriers; and adjacent sub-carriers are orthogonally phased to each other. Each of

15   the sub-carriers is then modulated by a low rate data stream. As the sub-carriers are packed more closely than, for example, in frequency division multiplexing (FDMA), OFDM allows the frequency spectrum to be used more efficiently. In addition, OFDM does not require complex time switching, as in time division multiplexing (TDMA), and therefore does not suffer the overhead associated with time switching methods.

20   FIG. 1 shows an OFDM system for transmitting and receiving information. A serial stream of data symbols is provided via input 101 to a serial to parallel converter 102 that converts the single data stream to several parallel data streams. An inverse discrete Fourier transform (IDFT) module 104 processes the parallel data streams and produces a corresponding number of orthogonal modulated sub-carriers which are

25   provided to a parallel to serial converter 106. In response, the parallel to serial converter 106 provides a serial data signal to a cyclic prefix adder 108, and the cyclic prefix adder 108 produces a transmit data signal at output 110, and the transmit data signal is transmitted on a communication channel.

With further reference to FIG. 1, a corresponding received data signal on the

30   communication channel is provided to a cyclic prefix remover 114 via input 112. The cyclic prefix remover 114 removes the cyclic prefix from the received data signal and outputs a single stream of data to a serial to parallel converter 116. Resultant signals from the outputs of the serial to parallel converter 116 are provided to a discrete Fourier transform (DFT) module 118, which provides a corresponding plurality of demodulated

35   data streams to a parallel to serial converter 120. A serial data signal is then provided by the parallel to serial converter 120 to an equalizer 135 and to a channel estimator 130,

and the equalizer operates with the channel estimator 130 to determine the originally transmitted data from received data, and provide the received data via an output 135.

The cyclic prefix is employed to address distortion in the communication channel. Adding the cyclic prefix comprises repeating the last few samples of each data symbol at its beginning, prior to its transmission. The length of the cyclic prefix should be chosen to be greater than or equal to the duration of the impulse response of the communication channel. This allows equalization of the channel distortion in the frequency domain by using a single tap scalar equalizer for each carrier, independently. However, in order to do this the response of the communication channel needs to be characterized. In practice only an estimate of the communication channel's characteristics is used, hence the need for a channel estimator.

There are several methods of performing channel estimation, these include the following schemes; Pilot Symbol Assisted Modulation (PSAM), Blind Channel estimation, and a coded pilot method. Each of these is briefly described below.

PSAM adds periodic transmissions of known symbols or pilots. Pilots comprise data that is known by both the transmitter and the receiver. Therefore, communicating pilot symbols allows the receiver to determine the difference between what was transmitted and what was received, and thus compensate for any variations in the received symbols that are caused by transmission between the transmitter and the receiver i.e. the communication channel. An estimate of the characteristics of the communication channel is required to provide such compensation across time and frequency domains of the communication channel. When the time and frequency characteristics of the communication channel are varying rapidly, as in mobile communication applications for example, channel estimation must be performed more frequently, hence the need for more pilots to be transmitted in order to maintain reliable communication. Thus, reducing the available bandwidth for data transmission.

FIG. 2 shows a PSAM scheme graphically, where both pilot symbols 202 and data symbols 204 are shown in a three dimensional grid across time 206 and frequency 208 axes, and where the vertical axis represents transmission power 210 of the data and pilot symbols. In PSAM, the pilot symbols 202 are inserted at intervals across time and frequency between the data symbols 204. Consequently, part of the signal energy and bandwidth of the communication channel is used for transmitting the pilot symbols 202. A received data signal in a PSAM scheme is passed through a 2-D Wiener filter, which essentially performs interpolation based on the statistics of the communication channel so as to estimate the characteristics of the communication channel between the pilot symbols 202, i.e. where the data symbols 204 are received. In this way, the data symbols 204 can be correctly recovered by taking the estimated time and frequency

characteristics into consideration to provide channel equalisation. One implementation of the PSAM scheme is in terrestrial transmission in digital video broadcasting (DVB-T).

5 A PSAM scheme provides good channel estimation even when applied to time variant channels. However, when the normalized maximum Doppler spread is high, caused by fast changes in the communication channel characteristics, the frequency at which pilot symbols are required increases in order to track such fast changes. This results in more bandwidth being required for pilot symbols, up to ten percent of the bandwidth of the communication channel, and leaving less of the bandwidth for data

10 traffic.

Blind channel estimation does not use pilots. Instead, the data symbols themselves are used to estimate the communication channel. Consequently, bandwidth of the communication channel is preserved. Several blind channel estimation schemes for OFDM are known, however, their tracking ability in a communication channel whose characteristics change or vary with time, Rayleigh fading time variant channels, for

15 example, have not been as good as that of the PSAM scheme.

The coded pilot method is described in US patent 5912876 by H'mimy where a main signal, comprising a quadrature amplitude modulated (QAM) version of a signal to be transmitted, and a pilot signal, are coded separately and transmitted as part of an

20 OFDM signal. When the OFDM signal is received, the main signal portion is detected and an estimation of the communication channel is determined from the detected coded pilot signal portion. Then the detected main signal and the estimation of the communication channel are used to estimate the signal that was transmitted. The coded pilot method is simple to implement, and the coding enhances the detection of the main

25 and pilot signals, in the consequent channel estimation process.

However, a transceiver using the coded pilot method is necessarily more complicated due to the coding in the transmitter, and detection of the codes in the receiver. In addition, a portion of the bandwidth of the communication channel needs to be allocated to support the transmission of the coded signals, thus reducing the usable

30 portion of a predetermined bandwidth.

Hence, there is a need for a channel estimation scheme that provides good performance in a communication channel having varying frequency and time characteristics, while preserving the usable bandwidth of the communication channel.

Brief Summary of the Invention

The present invention seeks to provide a method and an apparatus for semi-blind communication channel estimation, which overcomes, or at least reduces the

5   abovementioned problems of the prior art.

Accordingly, in one aspect, the present invention provides a multi-carrier communication system comprising:

a transmitter comprising:

a data input for receiving an unknown data signal for transmission;

10   a known data input for receiving a known data signal;

a signal power ratio input for receiving a power ratio signal indicating a ratio for combining the unknown data signal and the known data signal; and

a data combiner coupled to the data input, the known data input and the signal power ratio input, the data combiner for combining the unknown data signal and

15   the known data signal in accordance with the power ratio signal to produce a composite output signal comprising discrete data signals, wherein each discrete data signal comprises at least a portion of the unknown data signal and at least a portion of the known data signal combined in accordance with the power ratio signal, the data combiner having an output adapted to provide the composite output signal to a multi-

20   carrier transmitter, wherein the multi-carrier transmitter transmits a transmit signal on a communication channel, wherein the transmit signal includes the composite output signal; and

a receiver comprising:

a multi-carrier receiver for receiving a receive signal corresponding to the

25   transmit signal on the communication channel, and the multi-carrier receiver having an output for providing a corresponding composite signal, wherein the corresponding composite signal comprises corresponding discrete data signals, and the corresponding composite signal being shaped by at least one signal shaping characteristic of the communication channel;

30   a channel estimator having a known data input for receiving the known data signal, an input coupled to receive the corresponding composite signal, an input coupled to receive the power ratio signal, and an input for receiving at least one estimate of the unknown data signal, the channel estimator for estimating the at least one signal shaping characteristic of the communication channel from at least the corresponding composite

35   signal, the at least the portion of the known data signal of at least some of the corresponding discrete data signals, the power ratio signal and the at least one estimate

of the unknown data signal, and the channel estimator having an output for providing at least one estimated communication channel characterising signal; and

an equalizer coupled to receive the corresponding composite signal, the known data signal, the power ratio signal and the at least one estimated communication channel characterizing signal, the equalizer for configuring at least one of its signal shaping characteristics to compensate for the at least one signal shaping characteristic of the communication channel, the configured equalizer for shaping the corresponding composite signal accordingly, and the equalizer having an output for providing at least one subsequent estimate of the unknown data signal.

In another aspect the present invention provides a multi-carrier transmitting system comprising:

a data input for receiving an unknown data signal for transmission;

a known data input for receiving a known data signal;

a signal power ratio input for receiving a power ratio signal indicating a ratio for combining the unknown data signal and the known data signal; and

a data combiner coupled to the data input, the known data input and the signal power ratio input, the data combiner for combining the unknown data signal and the known data signal in accordance with the power ratio signal to produce a composite output signal comprising discrete data signals, wherein each discrete data signal comprises at least a portion of the unknown data signal and at least a portion of the known data signal combined in accordance with the power ratio signal, the data combiner having an output adapted to provide the composite output signal to a multi-carrier transmitter, wherein the multi-carrier transmitter transmits a transmit signal on a communication channel, wherein the transmit signal includes the composite output signal.

In yet another aspect the present invention provides a multi-carrier receiving system comprising:

a multi-carrier receiver for receiving a receive signal on a communication channel, where in the receive signal includes a composite signal, and the multi-carrier receiver having an output for providing the composite signal, wherein the composite signal comprises discrete data signals, and wherein each discrete data signal comprises at least a portion of an unknown data signal and at least a portion of a known data signal combined in accordance with a signal power ratio signal, the composite signal being shaped by at least one signal shaping characteristic of the communication channel;

a channel estimator having a known data input for receiving the known data signal, an input coupled to receive the composite signal, an input coupled to receive the power ratio signal, and an input for receiving at least one estimate of the unknown data

signal, the channel estimator for estimating the at least one signal shaping characteristic of the communication channel from at least the composite signal, the at least the portion of the known data signal of at least some of the discrete data signals, the power ratio signal and the at least one estimate of the unknown data signal, and the channel

5   estimator having an output for providing at least one estimated communication channel characterising signal; and

an equalizer coupled to receive the composite signal, the known data signal, the power ratio signal and the at least one estimated communication channel characterizing signal, the equalizer for configuring at least one of its signal shaping characteristics to

10   compensate for the at least one signal shaping characteristic of the communication channel, the configured equalizer for shaping the composite signal accordingly, and the equalizer having an output for providing at least one subsequent estimate of the unknown data signal.

In still another aspect the present invention provides a method for determining

15   received data in a multi-carrier communication system, wherein a received signal includes a composite signal received on a communication channel having transmission characteristics, wherein the composite signal comprises a plurality of discrete data signals spaced in time and frequency, and wherein each discrete data signal comprises a data portion and a pilot portion, wherein the data portion comprises one of a

20   predetermined group of symbols, the method comprising the steps of:

a)   defining a set of the plurality of discrete data signals;

b)   setting a predetermined number of iterations;

c)   receiving the composite signal on the communication channel;

d)   selecting one of the plurality of discrete data signals to be estimated;

25   e)   selecting a group of the plurality of discrete data signals, wherein each discrete data signal of the group of the plurality of discrete data signals is relevant to determining the one of the plurality of data signals;

f)   normalising the group of the plurality of discrete data signals using at least the pilot portion of the one of the plurality of discrete data signals;

30   g)   estimating the transmission characteristics of the communication channel using the normalised group of the plurality of discrete data signals in step (f); and

h)   estimating the data portion of the one of the plurality of discrete data signals using the communication channel having the transmission characteristics estimated in step(g) and at least the pilot portion of the one of the plurality of discrete data signals.

Brief Description of the Drawings

An embodiment of the present invention will now be more fully described, by way of example, with reference to the drawings of which:

FIG. 1 shows a prior art communication system;

FIG. 2 shows a graphical representation of data and pilot symbols communicated on a communication channel of the communication system in FIG. 1;

FIG. 3 shows a communication system in accordance with the present invention;

FIG. 4 shows a graphical representation of data and pilot symbols communication on a communication channel of the communication system in FIG. 3;

FIG. 5 shows a flow chart diagram of a process in a receiver of the communication system in FIG. 3;

FIG. 6 shows a graphical representation of the APSB scheme in the communication system in FIG. 3; and

FIG. 7 and 8 shows comparative simulation results of the performance of the communication system in FIG.3 and prior art communication systems.

Detail Description of the Drawings

The present invention combines the advantages of the PSAM and the blind schemes to produce a scheme where data and pilot symbols are combined prior to transmission, and separated when received at a receiver. In addition, separation of the pilot symbols at the receiver is accomplished by treating the data as noise and applying an iterative process to detect the data symbols. An embodiment of the present invention will now be described.

In FIG. 3 a communication system 300 that communicates data using an added pilot semi-blind (APSB) scheme, in accordance with the present invention, comprises a combiner 305 having a data input 307 for receiving a data signal 306, a pilot input 309 for receiving a pilot signal 308, and an input 311 for receiving a data–to-pilot power ratio (DPR) signal. The data signal 306 comprises a series of data symbols while the pilot signal comprises a series of pilot symbols. The DPR signal includes information on power levels of the data signal 306 and the pilot signal 308 that are to be combined.

The combiner 305 includes a data conditioner 312 that receives the data signal 306 and the DPR signal and provides a conditioned data signal having a power level as indicated by the DPR signal. Similarly, the combiner 305 includes a pilot conditioner 313 that receives the pilot signal 308 and the DPR signal, and provides a conditioned pilot signal having a power level as indicated by the DPR signal.

The combiner 305 also includes an adder 314 for combining the conditioned data signal from the data signal conditioner 312 and the conditioned pilot signal from the pilot signal conditioner 313, by combining a series of conditioned data symbols and a series of conditioned pilot symbols, and providing data-pilot signals 310 via an output

5   318 of the combiner 305. The data-pilot signals 310 comprises a series of discrete data signals, where each discrete data signal has a data portion and a pilot data portion, and where the power levels of the data portion and the pilot portion are in accordance with the DPR signal. The data portion represents at least one data symbol

The data-pilot signals 310 are then processed, in sequence, by the serial to parallel

10   converter 102, the IFFT 104, the parallel to serial converter 106, the cyclic prefix adder 108, as described earlier, and a transmit data-pilot signal from output 310 is transmitted on a communication channel.

A corresponding receive data-pilot signal 314 received from the communication channel at input 315, is processed sequentially by the cyclic prefix remover 114, the

15   serial to parallel converter 116, the discrete Fourier transform 118 and the parallel to serial converter 120, as described earlier. An output data-pilot signal 316 from output 317 of the parallel to serial converter 120 is then provided to an APSB equalizer 325 and to an APSB channel estimator 320. The APSB channel estimator 320 and the APSB channel equaliser 325, each include an input 319 for receiving the pilot signal 308, and

20   each include an input 318 for receiving the DPR signal. The pilot signal 308 and the DPR signal may be stored in a memory (not shown). The APSB equaliser 325 operates with the APSB channel estimator 320 to determine the originally transmitted data 306 from the data-pilot signal 316 using the pilot signal 308 and the DPR signal, and provides the originally transmitted data 306 via output 330 of the APSB equaliser 325.

25   This is accomplished with an iterative process where a channel estimate $\hat{h}(n)$ is made by the APSB channel estimator 320, and a data estimate is then made using the previously obtained channel estimate and provided via output 330. The output 330 is coupled to the APSB channel estimator 320 to provide the channel estimate for a subsequent iteration of channel estimation and data estimation. When a predetermined

30   number of iterations have been performed, the originally transmitted data 306 is determined and provided via output 330 of the APSB equaliser 325.

With additional reference to FIG. 4, which shows the APSB scheme graphically, symbols of the data-pilot signal 317 are shown as discrete signals 405 in the three dimensional grid across time 206 and frequency 208 axes, and where the vertical axis

35   represents transmission power 210. Each discrete signal 405 comprises two portions, a data portion 407 and a pilot portion 409. The proportions are defined by the DPR signal, which is, as defined earlier, the ratio of the data signal power to the pilot signal power.

There is an optimum DPR where the performance of the communication system 300 would be best.

In accordance with the present invention as described, the pilot is transmitted along with the data simultaneously and in the same frequency band, therefore

5 transmission of the pilot advantageously does not consume bandwidth of the communication channel.

The data-pilot signal 314 received from the communication channel contains information about the communication channel, and the pilot portion of the data-pilot signal is known. Then, by treating the data portion as noise, an initial estimation of the

10 communication channel can be determined, and a data estimate obtained, with the initial estimation of the communication channel, using a least squares approach. With partial knowledge of the characteristics of the communication channel via the channel estimates and the data estimates, further iterations of channel and data estimation are performed, thereby improving the accuracy of the estimation and allowing an accurate

15 estimate of the transmitted data to be made after a predetermined number of iterations.

With additional reference to FIG. 5, a process 500 for estimating the transmitted data starts 505 with selecting 510 a set of cells S_D from the received set of data-pilot cells S. Next the predetermined number of iterations is set 515, and a first cell, for example cell 430 at (n,l) is selected 535. Subsequently, cells relevant to the estimation

20 of the cell 430 are selected 540. Here, the selected cells are the eight cells adjacent to the cell 430 as typically the closest surrounding cells are selected. Using the time 206 n and frequency 208 l axes as references, the cell 430 is identified as cell (n,l) and the relevant selected cells as $S^P(n,l)$. All the relevant selected cells $S^P(n,l)$ are then normalised 542, for example by dividing the selected relevant cells by the pilot symbol

25 portion 409 to allow a suitable channel estimate to be made. An initial channel estimate 545 is then made using a 2-D Wiener filter, where the data portions 407 of the selected relevant cells are treated as noise, and the result is an estimate of the communication channel at (n,l). Subsequently, a data estimate 550 is made using the least squares approach, and a decision 555 is made as to which symbol was transmitted. The decision

30 step 555 may involve processes such as interleaving and coding before a decision is made.

A determination 560 is then made as to whether all the cells in the set S have been determined. When not all the cells in the set S have been determined, then the next cell of the set S of cells is selected 570, and the process 500 returns to step 540 of selecting

35 relevant cells to the newly selected cell. The looping back through step 570 continues until all the cells in the set S have been determined. When all the cells in the set S have been determined, the determination at step 560 is true, a counter (not shown) indicating

the number of iterations i is incremented 562, and a determination 565 is made as to whether the number of iteration i have reached the predetermined number of iterations set in step 515. When the number of iterations i has reached the predetermined number of iterations set in step 515, the process 500 ends 567. However, when the number of
5   iterations i has not reached the predetermined number, the process 500 returns to step 535 and repeats as described above.

Returning now to FIG. 3 a functional description of the communication system 300 now follows. The rules for notation that are employed are:

- in the time domain, signals at index (n,l) are written as $h(n,l)$, vectors as $h(n)$, and
10   matrices as $H(n)$;

- in the frequency domain a subscript F is added (e.g. $h_F(n,l)$, $h_F(n)$, $H_F(n)$); and sub-vectors/matrices are capped with a tilde e.g. $\tilde{h}$ or $\tilde{H}$, and estimated entities

- with a hat e.g. $\hat{h}$;

- superscripts ^{T}, ^{H} and * denote the operations matrix transpose, Hermitian and
15   conjugation, respectively, and E denotes expectation; and

- elements of vectors/matrices are denoted as $[h]_i / [H]_{i,j}$.

The elementary sample period of an OFDM system model is T, the number of subcarriers is L, the number of cyclic prefix samples is D, and the maximum number of channel response samples is

20
$$B+1 \ (\leq D)$$

while the total number of samples in one OFDM symbol is
$$P = L+D.$$
and, the OFDM symbol period is defined as
$$T_s = PT.$$

25   The complex baseband representation of the communication channel 605, particularly a mobile wireless communication channel, impulse response at time t is described by

$$h(t,\tau) = \sum_l \gamma_l(t) \, \delta(\tau - \tau_l(t)) \qquad\qquad (1)$$

30   where $\tau_l(t)$ and $\gamma_l(t)$ are the delay and complex amplitude of the lth path, respectively. The power delay profile of the channel is defined as

$$p(\tau) = E\left[h(t, \tau) \, h^*(t, \tau)\right] \qquad\qquad (2)$$

Assuming an exponentially decaying power delay profile with

$$p(\tau) = A \exp\left(-\tau/\tau_{rms}\right)$$

where τ_{rms} is a parameter of the channel known as the root mean squared delay spread, and A is a normalizing constant. The normalized τ_{rms} is given as τ_{rms}/T.

Due to the relative mobility of the transmitter and receiver, a maximum Doppler spread, f_D, will occur in the received signal. This is accounted for with a time-variant channel where a high f_D implies a fast varying channel. The normalized maximum Doppler spread is defined as $f_D T_s$, based on the assumption that the channel coefficients are time invariant over each OFDM symbol period T_s.

A function which is useful for analysis in OFDM is the time-variant transfer function obtained from the Fourier transform of equation (1) above, with respect to delay τ, which produces the equation below.

$$h_F(t,f) = \int_{-\infty}^{\infty} h(t,\tau) \exp(-j2\pi f\tau)\, d\tau \tag{3}$$

Assuming a wide sense stationary uncorrelated scattering (WSSUS), and a Rayleigh fading channel with Jakes' spectrum, the autocorrelation of the channel is separable in time t, and frequency f, and can be written as follows.

$$r_{h_F h_F}(t,t';f,f') = E\left[h_F(t,f)h_F^*(t',f')\right]$$
$$= r_t(\Delta t) r_f(\Delta f) \tag{4A}$$

where $\Delta t = t - t'$; and $\Delta f = f - f'$. We have

$$r_f(\Delta f) = \frac{1 - \exp(-DT(1/\tau_{rms} + j2\pi\Delta f))}{(1 - \exp(-DT/\tau_{rms}))(1 + j2\pi\tau_{rms})}$$

$$\approx \frac{1}{1 + j2\Delta f\tau_{rms}} \tag{4B}$$

assuming $[(D)/(\tau_{rms}/T)] \gg 1$, and

$$r_t(\Delta t) = J_0(2\pi F_D \Delta t) \tag{4C}$$

with $J_0(\cdot)$ being the zeroth order Bessel function of the first kind.

The discrete channel model of a communication channel of order $B(<D)$ can be described as $h(n,l) = h(t = nT_s, \tau = lT)$, assuming that $\tau_k(t)$ is uniformly spaced at intervals of T and that $p(\tau) = 0$ for $\tau > \tau_{max} = BT$ and represented as a vector at time n as
$$\mathbf{h}(n) = [h(n,0)...h(n,L-1)]^T$$
$$= [h(n,0)...h(n,B)0...0]^T \in C^{L\times 1} \tag{5}$$
where C is a complex number.

The channel coefficients in the frequency domain are obtained in a similar way
to equation (2), but in discrete time as

$$\mathbf{h}_F(n) = \sqrt{L}\mathbf{F}\mathbf{b}\quad (n) = \left[h_F(n,0)...h_F(n,L-1) \right]^T \qquad (6)$$

where $\mathbf{F}$ is the $L \times L$ unitary discrete Fourier transform (DFT) matrix with
$[\mathbf{F}]_{n,l} = \exp(-2\pi n l / L) / \sqrt{L}$, and $\mathbf{F}^H$ is the corresponding inverse DFT (IDFT) matrix.
The auto-correlation of the discrete channel, $r_{h_F h_F}(n,l)$, can be obtained from the
equations (4A), (4B) and (4C) by replacing $\Delta f = 1/(LT)$ and $\Delta t = nT_S$.

The received OFDM signal may be considered a digital signal in a 2-D cell
structure with indices $(n,l) \in S$.

where $S = \{(n,l) : 0 \leq n \leq N-1, 0 \leq l \leq L-1\}$, and where N is the total number of cells in
the time direction, n, and L is the number of sub-carriers as defined earlier.

A model for a 2-D Wiener filter will now be described, as such a filter is used to
form estimates from sampled signals at the receiver. Estimates will be obtained at
index $(n,l) \in S_D$ from the sampled signals at indices $(n',l') \in S_P^{(n,l)}$, where S_D and $S_P^{(n,l)}$
are subsets of S, and where $S_P = \cup_{n,l} S_P^{(n,l)}$. The definitions of S_D, S_P and $S_P^{(n,l)}$ will be
provided later.

In a 2-D wide sense stationary (WSS) stochastic process $\breve{h}_F(n,l)$
which contains information of the desired signal $h_F(n,l)$ and is corrupted by correlated
noise $h_F(n,l) v_F(n,l)$ and additive white Gaussian noise (AWGN), $u_F(n,l)$, as follows:

$$\breve{h}_F(n,l) = h_F(n,l) + h_F(n,l) v_F(n,l) + u_F(n,l) \qquad (7)$$

The notation $h_F(n,l)$ and $u_F(n,l)$ indicate that the channel frequency response
is estimated from its noisy samples. It is assumed that $v_F(n,l)$ and $u_F(n,l)$ are
white, and that $h_F(n,l), v_F(n,l)$ and $u_F(n,l)$ are mutually un-correlated, and zero mean
stochastic processes. In the prior art PSAM scheme, the correlated noise component is
absent, in contrast, for the APSB scheme of the present invention, as described, the
correlated noise component is advantageously reduced through successive iterative
measures.

For the 2-D Wiener filter, the estimator for $h_F(n,l)$ is given as:

$$\bar{h}_F(n,l) = \sum_{\{n',l'\} \in S_P^{(n,l)}} w(n,l;n'l') \breve{h}_F(n',l'), \qquad \forall (n,l) \in S_D \qquad (8)$$

where $w(n,l;n'l')$ is the weight applied on $\breve{h}_F(n',l')$ to estimate $h_F(n,l)$.

Based on the assumption that a total of Δ_N and Δ_L pilots are used in the n and l
direction, respectively, for each estimation of $h_F(n,l)$, a vector $\breve{\mathbf{h}}(n,l) \in C^{\Delta_N \Delta_L \times 1}$ is
performed by stacking the elements from $\breve{h}_F(n',l'), \forall (n',l') \in S_P^{(n,l)}$, and a vector
$\mathbf{w}(n,l) \in C^{\Delta_N \Delta_L \times 1}$ is formed by stacking the conjugate elements of $w(n,l;n'l')$.

Equation (8) can be re-written as follows.

$$\hat{h}_F(n,l) = \mathbf{w}^H(n,l)\,\breve{\mathbf{h}}_F(n,l) \tag{9}$$

To minimize the difference $h_F(n,l) - \hat{h}(n,l)$ in the mean square sense and obtain the optimum tap-weight vector, the orthogonality principle is now applied to equation (9), resulting in the equation below.

$$E\left[(h_F(n,l) - \hat{h}_F(n,l)).h_F^*(n'',l'')\right] = 0, \qquad \forall\{n'',l''\} \in S_P^{(n,l)} \tag{10}$$

Substituting equation (8) into equation (10), the Wiener-Hopf equation, with optimum tap weights $w_0(n,l;n',l')$, is obtained as provided below.

$$\sum_{\{n',l'\}\in S_P^{(n,l)}} w_0(n,l;n',l')E\left[\breve{h}_F(n',l')\breve{h}_F^*(n'',l'')\right] = E\left[h_F(n,l)h_F^*(n'',l'')\right] \tag{11}$$

Defining the cross-correlation of $h_F(n,l)$ and $\breve{h}_F(n,l)$ and the autocorrelation of $\breve{h}_F(n,l)$, respectively as

$$r_{\breve{h}_F h_F}(n-n'',l-l'') = E\left[\breve{h}_F(n,l)h_F^*(n'',l'')\right] \tag{12}$$

$$r_{\breve{h}_F h_F}(n'-n'',l'-l'') = E\left[\breve{h}_F(n',l')h_F^*(n'',l'')\right] \tag{13}$$

Letting $\mathbf{r}_{\breve{h}_F h_F}(n,l) = E\left[\breve{\mathbf{h}}_F(n,l)h_F^*(n,l)\right] \in C^{\Delta_N \Delta_L \times 1}$ be the cross-correlation vector and $\mathbf{R}_{\breve{h}_F \breve{h}_F}(n,l) = E\left[\breve{\mathbf{h}}_F(n,l)\,\breve{\mathbf{h}}_F^H(n,l)\right] \in C^{\Delta_N \Delta_L \times \Delta_N \Delta_L}$ be the autocorrelation matrix $\breve{h}_F(n,l)$ formed from the elements of equations (12) and (13), respectively. Therefore, the Wiener-Hopf equation can be re-written in matrix notation as

$$\mathbf{w}_0(n,l) = \mathbf{R}_{\breve{h}_F \breve{h}_F}^{-1}\,\mathbf{r}_{\breve{h}_F h_F}(n,l) \tag{14}$$

Assuming that $h_F(n,l), v(n,l)$ and $u(n,l)$ are mutually un-correlated and wide-sense stationary white stochastic processes, using the definition provided by equation (7), and letting Δn and Δl be the discrete time and frequency difference indices, respectively, equations (12) and (13) can be written as follows.

$$r_{\breve{h}_F h_F}(\Delta n, \Delta l) = r_{h_F h_F}(\Delta n, \Delta l) \tag{15}$$

$$r_{\breve{h}_F \breve{h}_F}(\Delta n, \Delta l) = r_{h_F h_F}(\Delta n, \Delta l) + r_{h_F h_F}(\Delta n, \Delta l) \times r_{v_F v_F}(\Delta n, \Delta l)$$

$$+ r_{v_F v_F}(\Delta n, \Delta l)$$

$$= \begin{cases} r_{h_F h_F}(\Delta n, \Delta l) + \sigma_h^2 \sigma_v^2 + \sigma_u^2 & \text{for } \Delta n, \Delta l = 0 \\ r_{h_F h_F}(\Delta n, \Delta l) & \text{otherwise} \end{cases} \tag{16}$$

where $r_{h_F h_F}(\Delta n, \Delta l)$ is the autocorrelation matrix h_F.

$\mathbf{R}_{\tilde{h}_F \tilde{h}_F}$ and $\mathbf{r}_{\tilde{h}_F h_F}(n,l)$ can be simplified as

$$\mathbf{R}_{\tilde{h}_F \tilde{h}_F} = \mathbf{R}_{h_F h_F} + diag(\mathbf{R}_{h_F h_F})\sigma_v^2 + \sigma_u^2 \mathbf{I}$$

$$= \mathbf{R}_{h_F h_F} + (\sigma_h^2 \sigma_v^2 + \sigma_u^2)\mathbf{I} \qquad (17)$$

$$\mathbf{r}_{\tilde{h}_F h_F}(n,l) = \mathbf{r}_{h_F h_F}(n,l) \qquad (18)$$

And, where $\mathbf{R}_{h_F h_F}$ is the *NL* by *NL* autocorrelation matrix of h_F, then $\mathbf{R}_{h_F h_F}$ can be obtained from equation (4). If $\mathbf{R}_{h_F h_F}$, σ_v^2 and σ_u^2 are known, the optimum weight is

$$\mathbf{w}_0(n,l) = \left[\mathbf{R}_{h_F h_F} + (\sigma_h^2 \sigma_v^2 + \sigma_u^2)\mathbf{I}\right]^{-1} \mathbf{r}_{h_F h_F}(n,l) \qquad (19)$$

The mean square error (MSE) can then be obtained as

$$E\left[\left|h_F(n,l) - \hat{h}_F(n,l)\right|^2\right] \forall (n,l) \in S_D \qquad (20)$$

The minimum MSE (MMSE) at index *(n,l)* is then obtained by substituting equation (9) into equation (20), and using the optimum weight factor in equation (19),

$$MMSE(n,l) = \sigma_h^2 - \mathbf{w}_0^H(n,l)\mathbf{r}_{h_F h_F}(n,l) \qquad (21)$$

For simplicity, $\sigma_h^2 = 1$ for all subsequent equations.

Now letting the transmitted data in the frequency domain be

$d_F(n,l), \forall (n,l) \in S_D$, where *n* is the discrete time index and *l* the discrete frequency index. In an OFDM system, *n* refers to the OFDM block index, while *l* refers to the sub-carrier index, and pilots are denoted as $p_F(n,l), \forall (n,l) \in S_p$. It is assumed that $p_F(n,l)$ is deterministic and selected from a fixed set of alphabets, while $d_F(n,l)$, the data is a zero mean stochastic process. The variance of $p_F(n,l)$ is denoted as $\varepsilon^2 \sigma_d^2$, and in practical implementations $\varepsilon^2 \langle\langle 1$.

In accordance with the present invention, as described herein, pilot signals and data signals are added together and co-exist at all time, *n*, and frequency points , *l*, in order to conserve bandwidth, that is, $S_D = S_p = S$. The signal after the data signal and the pilot signal are combined is defined below.

$$x_F(n,l) = d_F(n,l) + p_f(n,l), \qquad \forall (n,l) \in S \qquad (22)$$

and $\sigma_x^2 = (1 + \varepsilon^2)\sigma_d^2$. Other definitions follow.

$$Data - to - Pilot\, Power\, Ratio\, (DPR) = \sigma_d^2/(\varepsilon^2 \sigma_d^2) = 1/\varepsilon^2$$

$$Signal - to - Pilot\, Power\, Ratio\, (SPR) = \sigma_x^2/(\varepsilon^2 \sigma_d^2) = 1 + 1/\varepsilon^2 = 1 + DPR$$

Assuming that the orthogonality of the OFDM system 600 is maintained, the signal after OFDM demodulation is

$$y_F(n,l) = h_F(n,l)x_F(n,l) + b_F(n,l)$$

$$= h_F(n,l)(d_F(n,l) + p_F(n,l)) + b_F(n,l) \qquad (23)$$

where $b_F(n,l)$ is an additive Gaussian noise (AWGN) stochastic process.

In accordance with the present invention, as described herein, channel-data estimation from the received signal $y_F(n,l)$ is performed in an iterative manner. For ease of description, the first iteration and subsequent iterations will be described separately.

FIRST ITERATION

Channel estimator 320 is a 2-D Wiener filter, as characterized earlier, estimates the channel response. First, the received signal $y_F(n,l)$ in equation (23) is normalized in the first iteration, thus.

$$\breve{h}_F(n,l) = y_F(n,l)/p_F(n,l)$$
$$= h_F(n,l) + h_F(n,l)d_F(n,l)/p_F(n,l) + b_F(n,l)/p_F(n,l)$$
$$= h_F(n,l) + h_F(n,l)v_{F1}(n,l) + u_{F1}(n,l) \qquad (24)$$

where $v_{F1}(n,l) = d_F(n,l)/p_F(n,l)$ is a data-dependent noise introduced due to the addition of the pilots to the data, and $u_{F1}(n,l) = b_F(n,l)/p_F(n,l)$ is AWGN. The subscript "1" indicates that the notation is specific to the first iteration. Similar use of subscript "i" will be employed for the ith iteration.

Assuming that the pilots, $p_F(n,l)$, are selected from a set of constant modulus symbols, we obtain

$$\sigma_{v1}^2 = E\left[\left|\frac{d_F(n,l)}{p_F(n,l)}\right|^2\right] = \frac{\sigma_d^2}{\varepsilon^2\sigma_d^2} = DPR \qquad (25)$$

and similarly

$$\sigma_{u1}^2 = \frac{\sigma_b^2}{\varepsilon^2\sigma_d^2} = \frac{1+\varepsilon^2}{\varepsilon^2}\frac{\sigma_b^2}{\sigma_x^2} = \frac{SPR}{SNR} \qquad (26)$$

where, as defined earlier, $SPR = (1+\varepsilon^2)/\varepsilon^2$, and $SNR = \sigma_x^2/\sigma_b^2$.

Equation (24) is in a similar form as equation (7) and the optimum tap weights can be obtained in a similar form as in equation (14) using time Δ_N and frequency Δ_L sampled signals. In this case, the autocorrelation matrix $\mathbf{R}_{\breve{h}_F\breve{h}_F}$, is independent of (n,l), however the cross correlation vector $\mathbf{r}_{\breve{h}_F h_F}(n,l)$ is usually dependent of (n,l).

FIG. 6 shows a graphical representation in the time/frequency domain, and $S_P^{(n,l)}$ is selected such that (n,l) is in the center of $S_P^{(n,l)}$. The set of indices encompased by $S_P^{(n,l)}$ will change, and may be visualized as a sliding window, as (n,l) changes. The relative index of (n,l) and $S_P^{(n,l)}$ however, will not change. Assuming wide sense stationarity, $\mathbf{R}_{\breve{h}_F h_F}(n,l)$ will also not change. Next, for example, the cross-correlation vector $\mathbf{r}_{\breve{h}_F h_F}(n,l)$ when estimating index 601 and 602 would be the same. When (n,l) is near to the boundary of S_P, such as when estimating index 603, the relative indices of (n,l) and $S_p^{(n,l)}$ would be different. Thus, $\mathbf{r}_{\breve{h}_F h_F}(n,l)$ would be different in this case.

Now, the MMSE estimator for the first iteration follows.

$$\bar{h}_{F1}(n,l) = \mathbf{W}_{0,1}^{H}(n,l)\, \breve{\mathbf{h}}_{F1}(n,l) \tag{27}$$

Thus, the first estimate of $x_F(n,l)$ is

$$\hat{x}_{F1}(n,l) = \frac{y_F(n,l)}{\bar{h}_{F1}(n,l)} \tag{28}$$

Then, in view of equation (22), the estimate for $d_F(n,l)$ is obtained as

$$
\begin{aligned}
d_{F1}(n,l) &= \hat{x}_{F1}(n,l) - p_F(n,l)\\
&= \frac{y_F(n,l)}{\bar{h}_{F1}(n,l)} - p_F(n,l)\\
&= \frac{h_F(n,l)(d_F(n,l) + p_F(n,l)) + b_F(n,l)}{\hat{h}_{F1}(n,l)} - p_F(n,l)
\end{aligned} \tag{29}
$$

If the channel estimation is perfect, i.e. $\hat{h}_{F1}(n,l) = h_F(n,l)$, then equation (26) becomes

$$\tilde{d}_{F1}(n,l) = d_F(n,l) + b_F(n,l)/h_F(n,l) \tag{30}$$

Then, a decision device or slicer, as is known in the art, is used on $\tilde{d}_{F1}(n,l)$ to obtain an estimate $\hat{d}_{F1}(n,l)$ of $d_F(n,l)$.

SUBSEQUENT ITERATIONS

For the second iteration, the normalization is carried out using $\hat{d}_{F1}(n,l) + p_F(n,l)$.

The normalization tries to remove the data-dependent noise $v_{F1}(n,l)$ that appears in the first iteration as characterized in equation (24). The normalization for the second iteration is carried out as follows:

$$
\begin{aligned}
\breve{h}_{F2}(n,l) &= \frac{y_F(n,l)}{\hat{d}_{F1}(n,l) + p_F(n,l)}\\
&= \frac{h_F(n,l)(d_F(n,l) + p_F(n,l)) + b_F(n,l)}{\hat{d}_{F1}(n,l) + p_F(n,l)}
\end{aligned} \tag{31}
$$

Making the substitution $\hat{d}_{F1}(n,l) = d_F(n,l)$ which is a good approximation when the probability of symbol error is small, equation (31) becomes

$$
\begin{aligned}
\breve{h}_{F2}(n,l) &= h_F(n,l) + \frac{b_F(n,l)}{\hat{d}_{F1}(n,l) + p_F(n,l)}\\
&= h_F(n,l) + u_{F2}(n,l)
\end{aligned} \tag{32}
$$

where

$$u_{F2}(n,l) = \frac{b_F(n,l)}{(\hat{d}_{F1}(n,l) + p_F(n,l))}$$

Equation (32) has the form of equation (7) with $\sigma_{v2}^2=0$ since $v_{F2}(n,l)=0$.

Then based on the assumption that $\varepsilon^2 \langle\langle 1, d_F(n,l)$ is modulated by quadrature phase shift keying (QPSK), and $p_F(n,l)\in\{\varepsilon\sigma_d(\pm1\pm j)/\sqrt{2}\}$, a QPSK constellation $|\hat{d}_{F1}(n,l)+p_F(n,l)|^2$ can have four possible values, each being equally likely to occur: $\sigma_d^2|(1+j)+\varepsilon(\pm1\pm j)|^2/2$. It should be noted that only the first quadrant of the QPSK signal constellation has been considered due to its 2-D symmetry. Hence, the required variance reduces to

$$\sigma_{u2}^2=\sigma_b^2\,E\left[\left|\hat{d}_{F1}(n,l)+p_F(n,l)\right|^{-2}\right]$$

$$=\sigma_b^2\frac{1+\varepsilon^4}{(1-\varepsilon^2)^2(1+\varepsilon^2)}\sigma_d^{-2}$$

$$=(1+\varepsilon^4)\frac{1}{(1-2\varepsilon^2+\varepsilon^4)}\frac{\sigma_b^2}{(1+\varepsilon^2)\sigma_d^2}$$

$$=(1+\varepsilon^4)(1+2\varepsilon^2+\cdots)SNR^{-1}$$

$$\approx(1+2\varepsilon^2)SNR^{-1} \qquad\qquad (33)$$

This approximation can be made when $\varepsilon^2\langle\langle 1$.

The 2-D Wiener filtering is applied accordingly in equation (9) and (14) by setting by setting $\sigma_v^2=\sigma_{v2}^2=0$ (i.e. there is no correlated noise component) and $\sigma_u^2=\sigma_{u2}^2$ Then the channel estimator is

$$\tilde{h}_{F2}(n,l)=\mathbf{w}_{0,2}^H(n,l)\,\breve{\mathbf{h}}_{F2}(n,l) \qquad\qquad (34)$$

Finally, the estimate $\hat{d}_{F2}(n,l)$ is obtained by following the steps as in equations (28) and (29).

The subsequent iterations can be extended from the second iteration. However, the assumption $\hat{d}_{F2}(n,l)=d_F(n,l)$ made in equation (32) will be more accurate in subsequent iterations and will therefore result in a better estimate for $\tilde{h}_{F3}(n,l),\hat{h}_{F4}(n,l)$, etc. As more iterations are performed, the channel estimate converges to the actual channel response.

In the prior art PSAM scheme, for every one pilot cell, there are on average $(\delta_N\delta_L-1)$ data cells. The effective average signal power after pilot insertion is

$$\sigma_x^2=\sigma_d^2\left[\frac{\eta^2+(\delta_L\delta_N-1)}{\delta_L\delta_N}\right] \qquad\qquad (35)$$

where $\eta^2\geq 1$ is the ratio of the power of the pilot to that of the data for PSAM scheme.

The signal-to-noise ratio (SNR) of the PSAM scheme can thus be

expressed as

$$SNR = 10\log_{10}(\sigma_x^2 / \sigma_b^2) = SNR' + SNR_{loss}(dB) \qquad (36)$$

$SNR' = \sigma_d^2 / \sigma_b^2$ is the SNR of the system without the presence of pilots, while SNR_{loss}

5   represents the additional SNR required to compensate for the loss of power which has

been allocated to the pilot. Therefore, it follows from equations (35) and (36) that

$$SNR_{loss} = 10\log_{10}(\sigma_x^2 / \sigma_d^2)\,dB$$
$$= 10\log_{10}\left(\frac{\eta^2 + (\delta_L\delta_N - 1)}{\delta_L\delta_N}\right)dB \qquad (37)$$

10

Since part of the bandwidth is used for pilot transmission, there is a reduction in

the effective data rate. The percentage bandwidth loss is

$$W_{loss} = \frac{1}{\delta_L\delta_N} \times 100\% \qquad (38)$$

15   When power boosted pilots (i.e. η)1) are used, we note that $SNR_{loss} > 0$ dB. The

actual SNR would have to be adjusted according to equation (36) for proper comparison

with other schemes. In the case when $\eta = 1$, however, no additional power is incurred

under the definition provided by equation (37). Regardless of the value of η, a

bandwidth loss is still incurred as seen from equation (38).

20   For the APSB scheme, in accordance with the present invention, as described,

the SNR is defined in a similar way as in equation (36), however with the APSB

scheme

$$SNR_{loss} = 10\log_{10}(\sigma_x^2 / \sigma_d^2)\,dB$$
$$= 10\log_{10}\left(\frac{SPR}{SPR-1}\right)dB \qquad (39)$$

25

In contrast to the prior art PSAM scheme, the APSB scheme SNR_{loss} is always

greater than 0 dB since ε)0. On the other hand, although the prior art PSAM scheme

suffers from bandwidth loss, the APSB scheme advantageously has $W_{loss} = 0$.

Both the prior art PSAM and the APSB schemes require sufficient OFDM

30   symbols to be received before an optimum channel estimate can be obtained.

Depending on the selection of the indices of the sampled signal, $S_P^{(n,l)}$, different

processing delays would be incurred in order for an optimum estimate to be made.

When a non-causal 2-D Wiener filter is used, such that the index (n,l) is at the center of $S_P^{(n,l)}$, the processing delay is then $\delta_N(\Delta N-1)/2$ for the prior art PSAM scheme. For the APSB scheme, the processing delay is the same with $\delta_N = 1$.

The results of simulations will now be presented. For the simulations, the common system parameters were set as $L = 32$, $D = 8$, $B = 7$, $\tau_{rms}/T = 0.5$, $\tau_{max}/T = 8$ and SNR = 20dB. It was assumed that the SNR is known and that $\sigma_h^2 = 1$, and a QPSK signal constellation for the transmitted signal and the pilot signal was used. In addition, $S_P^{(n,l)}$ was selected such that the index (n,l) is in the center of the 2-D indices of $S_P^{(n,l)}$. This results in the best MSE performance. It is possible to estimate the channel at indices

$$n \le (\Delta_N -1)/2 \, or \, n \ge N-(\Delta_N +1)/2, \quad and \quad l \le (\Delta_L -1)/2 \, or \, l \ge L-(\Delta_L +1)/2$$

in such a manner where a centered-interpolation is carried out. For the other indices, an off-centered interpolation is required to estimate the channel, giving some degradation of the MSE.

With reference to FIG. 7, consider first, the bit error rate (BER) performance at a maximum normalized Doppler spread, $f_D T$, of 0.005. Setting $\Delta_N = \Delta_L = 15$, the BER for the APSB scheme using different ε(SPR) for the first to the fifth iterations. A high SPR means that low amount of power is used for pilot transmission while a small SPR means that low amount of power is used for data. Both extreme cases imply that $\hat{h}_F(n,l)$ will be badly estimated. Hence, for each iteration, there exist an optimum point whereby the SPR will minimize the BER. It is observed that as the number of iterations increases, the performance improves, but at a diminishing marginal amount.

For the prior art PSAM scheme, as we increase η, the channel estimation improves. However, since SNR_{loss} also increases, less power is allocated for the data for a given SNR. Thus, the improvement of the BER brought about by better channel estimation would be offset at some point as η increases. Consequently, there is an optimum η that minimizes the BER as well.

With both the prior art PSAM scheme and the APSB scheme, performance improves when larger filter taps are used i.e. larger Δ_N and Δ_L. For the PSAM scheme, the performance also improves for smaller Δ_N and Δ_L when closer sampling intervals are employed, while for the APSB scheme, more iterations (K) results in better performance, although the marginal performance gain would decrease for both cases. Hence, the system design is flexible since increased complexity of the PSAM scheme, and the APSB scheme would improve performance. However, the APSB scheme, in accordance with the present invention as described, advantageously does not suffer

bandwidth loss in contrast to the PSAM scheme which will always, have loss of bandwidth.

In general the BER is relatively robust to the selection of ε and η, and is consistent for other values of Δ_N ($= \Delta_L$), Δ_N ($= \Delta_L$), and different values of Doppler spread. Optimum values selected are as $\varepsilon_o = 0.43$ and $\eta_o = 4/3$ that minimize BER.

With reference to FIG. 8, assuming that the SNR is known, the graph shows a plot of BER vs SNR for the prior art switched based blind scheme 801, and PSAM 802 scheme, and the APSB scheme 803 and 804, using $\varepsilon_o = 0.43$ and $\eta_o = 4/3$, at $f_D T_s = 0.005$. It is observed that when $\Delta_N = \Delta_L = 15$, plot 803, the SNR of the of the APSB scheme is 1 dB worse than the PSAM scheme. However, the BER performance of the APSB scheme is superior to the switched based blind scheme.

The present invention, as described, provides an added pilot semi-blind scheme that does not consume bandwidth, and is suitable for use in a mobile communication system.

This is accomplished by adding data and pilot data at the transmitter prior to transmission on a communication channel, and using a 2-D Wiener filter to recover the transmission characteristics of the communication channel using the pilot data. Then through an iterative process of estimating the communication channel and estimating the data, the communication channel can be estimated and the transmitted data recovered.

The present invention provides a method and an apparatus for semi-blind communication channel estimation, which overcomes, or at least reduces the abovementioned problems of the prior art

It will be appreciated that although only one particular embodiment of the invention has been described in detail, various modifications and improvements can be made by a person skilled in the art without departing from the scope of the present invention.

We claim: